APPARATUS AND METHODS FOR INTERSYMBOL INTERFERENCE COMPENSATION IN SPREAD SPECTRUM COMMUNICATIONS

BACKGROUND OF THE INVENTION

The present invention relates to communications apparatus and methods, and more particularly, to spread spectrum communications apparatus and methods.

Wireless communications systems are widely used to communicate voice and other data, and the use of such systems is increasing through the development of new applications. For example, in addition to traditional voice telephony applications, wireless systems are increasingly being used to provide data communications services such as internet access and multimedia applications.

FIG. 1 illustrates a typical direct sequence spread spectrum (DS-SS) signal generator, as might be used in a code division multiple access (CDMA) communications system. A data sequence is spread by a spreading sequence, which typically has a much higher baud rate. The spread signal thus produced is passed through a pulse shaping filter to generate a baseband signal $s(t)$, which is given by:

$$s(t) = \sum_{i=-\infty}^{\infty} \alpha(i) f_i(t - iT), \tag{1}$$

$$f_i(t) = \sum_{l=0}^{N-1} a_i(l) p(t - lT_c), \tag{2}$$

where $f_i(t)$ is the spreading waveform for the ith symbol, $\alpha(i)$ is the ith data symbol, $a_i(l)$ is the lth "chip" of the spreading sequence in the ith symbol interval, N is the processing gain, T_c is the chip duration, $T = NT_c$ is the symbol duration, and $p(t)$ is the chip pulse. The baseband signal $s(t)$ is then typically modulated by a carrier signal, and the resultant data-modulated carrier signal is transmitted in a communications medium, e.g., in air, wireline or other medium.

The channel experienced by a transmitted wireless DS-SS signal is typically modeled as a dispersive channel with an impulse response of the form:

$$g(t) = \sum_{l=0}^{L-1} g_l \delta(t - \tau_l) \qquad (3)$$

where L is the number of multipaths, and g_l and τ_l are the complex-valued attenuation

5   factor and delay for the lth path, respectively. The baseband equivalent signal

received over such a channel can be expressed as:

$$y(t) = \sum_i \alpha(i) h_i(t - iT) + n(t), \qquad (4)$$

where:

$$h_i(t) = \sum_{l=0}^{L-1} g_l f_i(t - \tau_l), \qquad (5)$$

10   and $n(t)$ includes thermal noise and multi-user interference.

Conventionally, a RAKE receiver **200** as shown in FIG. 2 may be used to

recover information from a DS-SS signal. A radio processor **220** converts a received

signal received via an antenna **210** to baseband, including filtering the signal based on

the chip pulse shape and sampling the result. A RAKE processor **230** includes a

15   correlator **232** that correlates the sampled signal with a spreading sequence at a

plurality of offset correlation times. For example, the correlator may include J RAKE

"fingers," each matched to one signal ray $(J = L)$, and a correlation between the

received signal and a delayed version of the spreading sequence may be calculated at

each finger. A combiner **234** typically employs maximum ratio combining (MRC) to

20   combine the correlation values produced by the correlator **232**, typically based on

channel coefficient estimates produced by a channel estimator **240**. Channel delay

estimates generated by the channel estimator **240** may be used to determine the offset

correlation times used by the correlator **232**.

One important feature of so-called "third generation" wireless

25   communications systems is the ability to provide services with a wide range of data

rates to meet the varying information transmission needs of various services such as

voice and data. For example, in IS-2000 and wideband CDMA (W-CDMA) wireless

communications systems, multiple data rates may be achieved by using various

combinations of codes, carriers and/or spreading factors. More particularly, in W-CDMA systems, the spreading factors of physical channels may range from 256 to 4, providing corresponding data rates from 15K baud per second (bps) and 0.96 Mbps.

5   For a physical channel employing a low spreading factor, a conventional RAKE receiver may not perform well if the channel is dispersive. This performance degradation may arise because the processing gain provided by signal spreading may not be sufficient to reject inter-symbol interference (ISI) arising from multipath propagation. Consequently, user throughput and coverage may be limited by multipath delay spread.

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SUMMARY OF THE INVENTION

According to embodiments of the present invention, a communications signal representing symbols encoded according to respective portions of a spreading sequence is decoded. Time-offset correlations of the communications signal with the 15   spreading sequence are generated. The time-offset correlations are combined to generate first estimates for the symbols. Intersymbol interference factors that include a relationship among different portions of the spreading sequence are determined. A second estimate for one of the symbols is generated from the first estimates based on the determined intersymbol interference factors.

20   An intersymbol interference factor may include a relationship between a first portion of the spreading sequence associated with the one symbol and a second portion of the spreading sequence associated with another symbol. An intersymbol interference factor may be determined, for example, from the spreading sequence and a channel estimate for a channel over which the communications signal is 25   communicated. The second estimate may be generated from the first estimates using, for example, a sequence estimation procedure that employs a branch metric that is a function of the determined intersymbol interference factors. Alternatively, a linear equalization procedure that uses weighting factors generated based on knowledge of the symbol dependence of the spreading sequence may be used.

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According to other embodiments of the present invention, a communications signal representing symbols encoded according to respective portions of a spreading sequence is decoded. A plurality of time-offset correlations of the communications signal with the spreading sequence is generated. The plurality of time-offset

5   correlations are combined to generate a first estimate for one of the symbols. An intersymbol interference factor that includes a relationship among different portions of the spreading sequence is determined. A second estimate for the one symbol is generated from the first estimate based on the determined intersymbol interference factor.

10   According to yet other embodiments of the present invention, a communications signal representing symbols encoded according to a spreading sequence is decoded. Time time-offset correlations of the communications signal with the spreading sequence are generated. Weighting factors are generated from a channel estimate for a channel over which the communications signal is

15   communicated and knowledge of an interfering component of the communications signal. The time-offset correlations are combined according to the determined weighting factors to generate first estimates of the symbols. Intersymbol interference factors are determined from the spreading sequence, and a second estimate for one of the symbols is generated from the first estimates based on the determined intersymbol

20   interference factor.

The present invention may be embodied as methods and apparatus. For example, the present invention may be embodied in a receiver included in a communications apparatus, such as a wireless terminal, wireless base station, or other wireless, wireline or optical communications apparatus.

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BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a schematic diagram illustrating a conventional direct sequence spread spectrum (DS-SS) transmitter.

FIG. 2 is a schematic diagram illustrating a conventional DS-SS receiver.

FIG. 3 is a schematic diagram illustrating a signal processing apparatus according to embodiments of the present invention.

FIG. 4 is a schematic diagram illustrating a RAKE receiver according to embodiments of the present invention.

FIG. 5 is a flowchart illustrating exemplary operations for generating a symbol estimate according to embodiments of the present invention.

FIG. 6 is a flowchart illustrating exemplary operations for generating an intersymbol interference (ISI) factor according to embodiments of the present invention.

FIGs. 7 and 8 are charts graphically illustrating signal constellation partitioning for a reduced state sequence estimation (RSSE) process according to embodiments of the present invention.

FIG. 9 is a schematic diagram illustrating a generalized RAKE (G-RAKE) receiver according to still other embodiments of the present invention.

FIG. 10 is a flowchart illustrating exemplary operations for determining an ISI factor according to embodiments of the present invention.

FIG. 11 is a schematic diagram illustrating a receiver according to yet other embodiments of the present invention.

FIG. 12 is a chart illustrating potential performance of a conventional receiver in comparison to potential performance of a receiver according to embodiments of the present invention.

DETAILED DESCRIPTION

The present invention now will be described more fully hereinafter with reference to the accompanying drawings, in which preferred embodiments of the invention are shown. This invention may, however, be embodied in many different forms and should not be construed as limited to the embodiments set forth herein; rather, these embodiments are provided so that this disclosure will be thorough and complete, and will fully convey the scope of the invention to those skilled in the art. In the drawings, like numbers refer to like elements throughout.

In the present application, FIGs. 3-11 are schematic diagrams, flowcharts and signal constellation diagrams illustrating exemplary communications apparatus and operations according to embodiments of the present invention. It will be understood that blocks of the schematic diagrams and flowcharts, and combinations of blocks therein, may be implemented using one or more electronic circuits, such as circuits included in a wireless terminal or in a wireless communications system (*e.g.*, in a cellular base station or other device), or circuitry used in other types of wireless, wireline, optical and other communications systems. It will also be appreciated that, in general, blocks of the schematic diagrams and flowcharts, and combinations of blocks therein, may be implemented in one or more electronic circuits, such as in one or more discrete electronic components, one or more integrated circuits (ICs) and/or one or more application specific integrated circuits (ASICs), as well as by computer program instructions which may be executed by a computer or other data processing apparatus, such as a microprocessor or digital signal processor (DSP), to produce a machine such that the instructions which execute on the computer or other programmable data processing apparatus create electronic circuits or other means that implement the functions specified in the block or blocks. The computer program instructions may also be executed on a computer or other data processing apparatus to cause a series of operations to be performed on the computer or other programmable apparatus to produce a computer implemented process such that the instructions which execute on the computer or other programmable apparatus provide operations for implementing the functions specified in the block or blocks. Accordingly, blocks of the schematic diagrams and flowcharts support electronic circuits and other means that perform the specified functions, as well as operations for performing the specified functions.

It will also be appreciated that the apparatus and operations illustrated in FIGs. 3-11 may be implemented in a variety of communications environments, including wireless, wireline and optical communications environments. For example, the communications apparatus and operations illustrated in FIGs. 3-11 may be embodied in a wireless terminal, a wireless base station, a wireline communications device, an

optical communications device, or other communications apparatus. It will be appreciated that the processing apparatus and operations illustrated in FIGs. 3-11 may be combined with other apparatus and operations (not shown), including additional signal processing apparatus (*e.g.*, circuits that provide such functions) and operations.

5   According to some embodiments of the present invention, a communications signal representing a symbol encoded according to a spreading sequence is decoded by generating time-offset correlations of the communications signal and the spreading sequence, and combining the correlations to generate a first estimate of the symbol, *e.g.*, as might be done in a RAKE processor or a modified RAKE processor. This first 10 estimate is revised using an estimation procedure, such as a maximum likelihood sequence estimation (MLSE) procedure, a decision feedback sequence estimation (DFSE) procedure or a reduced state sequence estimation (RSSE) procedure, that uses intersymbol interference (ISI) factors that relate portions of the spreading sequence, *e.g.*, ISI factors generated from channel estimates and cross-correlations of the 15 spreading sequence. For example, the sequence estimation procedure may use a branch metric that is a function of ISI factors.

FIG. 3 illustrates an apparatus **300,** according to embodiments of the present invention, for decoding a communications signal **301** that represents a symbol sequence encoded according to a spreading sequence. A correlator **310** generates 20 time offset correlations **315** of the communications signal **301** with a spreading sequence **303**. A combiner **320,** *e.g.*, a RAKE combiner, combines the time-offset correlations **315** to generate first estimates **325,** *e.g.*, decision statistics, for symbols. A symbol estimator **340** generates second estimates **345** for symbols from the first estimates **325** based on ISI factors **335** generated by an ISI factor determiner **330.** 25 The ISI factors **335** include a relationship between portions of the spreading sequence, which may be generated, for example, responsive to a channel estimate **302** and the spreading sequence **303** as described in greater detail below.

According to some embodiments of the present invention, a sequence estimation procedure that employs a branch metric that is a function of an ISI factor is 30 used to revise symbol estimates produced by a RAKE processor. Two structures used

in maximum likelihood sequence estimation (MLSE) procedures are the Forney form and the Ungerboeck form, as described in G.D. Forney, "Maximum-Likelihood Sequence Estimation of Digital Sequences in the Presence of the Intersymbol Interference," *IEEE Trans. Inform. Theory*, vol. IT-18, no. 5, pp. 363-378 (May 1972)

5 and G. Ungerboeck, "Adaptive Maximum Likelihood Receiver for Carrier Modulated Data Transmission Systems," *IEEE Trans. Commun.*, vol. COM-22, no. 3, pp. 624-635 (March 1974), respectively. Each form typically employs the well-known Viterbi algorithm. Typically, the branch metrics used in the Viterbi algorithms for the Forney and Ungerboeck forms are different. If the Forney form is used, the branch metric

10 typically is an Euclidean metric, whereas, in the Ungerboeck form, the branch metric is typically the Ungerboeck metric. A Forney form receiver also typically uses a whitening filter and a discrete matched filter, both of which generally depend on the signal waveform.

In CDMA systems, the scrambling spreading sequence applied to a symbol

15 sequence to be transmitted often varies from symbol to symbol, *i.e.*, the scrambling sequence has a period greater than the symbol period, such that successive symbols are spread according to different portions of the scrambling sequence. If a Forney form were used in a receiver for a signal spread in such a symbol-dependent manner, the whitening filter and discrete matched filter used in the received would generally

20 need to change from symbol to symbol, making the Forney form less attractive for use in decoding such signals.

According to some embodiments of the present invention, an Ungerboeck form is used. The branch metric at the ith stage of the Viterbi decoder used in an MLSE procedure may be given by:

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$$M_H(i) = \mathrm{Re}\left\{a_i^*\left[2z(i) - s_{0,i}\alpha_i - 2\sum_{\substack{l \\ l>0}} s_{l,i}\,\alpha_{i-l}\right]\right\} \tag{6}$$

where α_i is the ith hypothesized symbol along the trellis path, and

$$z(i) = \sum_{j=0}^{L-1} g_j^* \int_{-\infty}^{\infty} f_i^*(t - \tau_j) y(t) dt \tag{7}$$

$$s_{l,i} = \sum_{n=1-N}^{N-1} C_{i,i-l}(n)(\phi_g(t) * \phi_p(t))\big|_{t=lT-nT_c}. \tag{8}$$

5   In the above equations, the parameter $z(i)$ is the output of a RAKE processor, $s_{l,i}$ is an intersymbol interference (ISI) factor (a so-called "s-parameter"), and $C_{i,i-l}(n)$, $\phi_g(t)$ and $\phi_p(t)$ are, respectively, the autocorrelation functions of the spreading sequence, channel impulse response $g(t)$, and chip pulse shape function $p(t)$. Furthermore:

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$$C_{i,j}(m) = \begin{cases} \sum_{l=0}^{N-1-m} a_i^*(l+m)a_j(l), & 0 \le m \le N-1 \\ \sum_{l=0}^{N-1+m} a_i^*(l)a_j(l-m), & 1-N \le m < 0 \end{cases} \tag{9}$$

$$\phi_p(t) = \int_{-\infty}^{\infty} p^*(\tau)p(t+\tau)d\tau \tag{10}$$

$$\phi_g(t) = \int_{-\infty}^{\infty} g^*(\tau)g(t+\tau)d\tau$$

$$= \sum_{j=0}^{L-1}\sum_{k=0}^{L-1} g_j^* g_k \delta(t + \tau_j - \tau_k). \tag{11}$$

15   Typically, the autocorrelation function of the pulse shape is nonzero only within a finite interval, such that:

$$\phi_p(t) \approx 0, \quad |t| > L_0 T_c. \tag{12}$$

Note that

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$$s_{l,i} \approx 0, \quad l > l_{max}, \tag{13}$$

for some l_{max} that depends on the pulse shape and delay spread.

FIG. 4 illustrates a receiver **400** according to embodiments of the present invention that uses an MLSE procedure that employs ISI factors, such as the s-

25   parameters described above, to revise symbol estimates produced by a RAKE

processor. An antenna **410** receives a communications signal **401**, which is processed by a radio processor **420** to generate a baseband signal **425**. A RAKE processor **430** includes a correlator **432** that generates time-offset correlations **433** of the baseband signal **425** with a spreading sequence **445** produced by a spreading sequence

5   generator **440**. The time-offset correlations **433** may be for correlation times corresponding to delays **455a** of a channel estimate **455** produced by a channel estimator **450**. A combiner **434** combines the time-offset correlations **433** according to channel coefficients **455b** of the channel estimate **455**, producing first estimates **435** of symbols represented by the communications signal **401**. An ISI factor

10   determiner **460** generates ISI factors **465** based on the channel estimate **455** and the spreading sequence **445**. A sequence estimator **470** generates second estimates **475** from the first estimates **435** based on the ISI factors **465**. For example, as described above with reference to equation (6), the sequence estimator **470** may process the first estimates **435** according to a sequence estimation procedure that uses a branch metric

15   that is a function of the ISI factors **465**.

According to other embodiments of the present invention, the number of states used in the sequence estimator **470** is varied responsive to the spreading factor, symbol modulation, and channel estimate (which, for purposes of the present application, may include the chip pulse shape function) for the channel over which a

20   received signal is communicated. In some embodiments, for example, for some l_{max} where $s_{l,i} \cong 0, l > l_{max}$, the number of states used in the sequence estimator **470** may be $A^{l_{max}}$, where A is the number of constellation points of the symbol modulation. When the nonzero -lag s-parameters are all of small magnitudes, the sequence estimator **470** may include a symbol-by-symbol detector. In still other embodiments, the value l_{max}

25   can be quantized to a finite set of values; consequently, the number of states used in the sequence estimator need only take values from a finite set of integer numbers.

In yet other embodiments of the present invention, the number of states used in the sequence estimator **470** is selected from a set consisting of 1 or A^L, where L is a predetermined number greater than zero, based on the delay spread (which, for

30   purposes of the present application, may be considered as part of the channel

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estimate) and spreading factor. In such a case, an appropriate branch metric is given by:

$$M_H(i) = \text{Re}\left\{\alpha_i^* \left[2z(i) - s_{0,i}\alpha_i\right]\right\}, \tag{14}$$

5

for the one state case, and

$$M_H(i) = \text{Re}\left\{\alpha_i^* \left[2z(i) - s_{0,i}\alpha_i - 2\sum_{l=1}^{L} s_{l,i}\alpha_{i-l}\right]\right\}, \tag{15}$$

for the A^L state case.

10   For the one state case, each symbol may be decided separately. Thus, one initial symbol estimate $z(i)$ can be used to determine the ith symbol. Under common operating conditions, the s-parameter $s_{0,i}$ is the same for all i and, accordingly, there is only one s-parameter.

It is common for forward error correction (FEC) decoding to follow symbol

15   estimation. Typical FEC decoders operate on so-called "soft" bit values, which can be viewed as a form of symbol estimation in which one of soft bit values constitute a symbol estimate. For the one state case discussed above, a soft value can be determined using the first symbol estimate $z(i)$ and the single s-parameter. For example, for a symbol corresponding to 3 bits, as in 8-PSK, a log-likelihood value

20   associated with each possible symbol value can be determined by taking the magnitude squared of the difference between $z(i)$ and $s_{0,0} \alpha_i$, where α_i corresponds to the possible symbol value. For a particular bit that makes up the 8-PSK symbol, four symbol values correspond to the bit being a "0" and four correspond to the bit being a "1". A technique for using such log-likelihood values to determine a soft value for a

25   bit is described in United States Patent Application Serial No. 09/587,995, entitled "Baseband processors and methods and systems for decoding a received signal having a transmitter or channel induced coupling between bits," to Bottomley et al., filed June 6, 2000 (Attorney Docket No. 8194-386). For the case of multiple states, standard techniques for extracting soft bit information for MLSE based sequence

detectors, such as the soft output Viterbi algorithm (SOVA) can be used. Such approaches are described in C. Nill and C. Sundberg, "List and soft symbol output Viterbi algorithms: extensions and comparisons," *IEEE Trans. Commun.*, vol. 43, pp. 277- 287, Feb./Mar./Apr. 1995, and in P. Hoeher, "Advances in soft-output

5   decoding," *Proc. Globecom '93*, Houston, TX, Nov. 29 - Dec. 2, pp. 793-797, 1993.

FIG. 5 illustrates exemplary operations **500**, according to embodiments of the present invention, for generating a symbol estimate using state number selection techniques, such as those described above. Time-offset correlations of a communications signal and a spreading sequence are generated (Block **510**). The

10   time-offset correlations are then combined to generate first estimates of symbols (Block **520**). ISI factors are determined (Block **530**). A number of states for a sequence estimation procedure is determined based on a channel estimate, spreading factor and symbol modulation (Block **540**) using, for example, one of the above-described procedures for selecting a number of sequence estimation states. A second

15   estimate of one of the symbols is generated from the first estimates using the determined number of states and a branch metric that is a function of the ISI factors (Block **550**).

FIG. 6 illustrates exemplary operations **600** for determining an ISI factor, in particular, an s-parameter, as described above with reference to equation (8). A

20   convolution of a channel impulse response autocorrelation function and a chip pulse shape autocorrelation function is determined (Block **610**). An aperiodic cross-correlation of the spreading sequence is determined (Block **620**). A convolution of these results is then calculated to generate an s-parameter (Block **630**).

As described above, the number of states used in the sequence estimator **470**

25   of FIG. 4 may depend on l_{max}. As l_{max} increases, however, the complexity of the sequence estimator **470** may increase to undesirable levels. According to other embodiments of the present invention, this complexity may be reduced by using a fixed number of states A^L. However, if $L << l_{max}$, this approach could result in significant performance degradation.

According to still other embodiments of the present invention, a tradeoff between complexity and performance may be achieved by using a form of decision-feedback sequence estimation (DFSE) in the sequence estimator **470** of FIG. 4. According to such an approach, $l_{max} + 1$ taps may be split into $l_F + 1$ feed-forward

5   taps and l_B feedback taps, where $l_F + l_B = l_{max}$. The decisions associated with the feedback taps are used in the branch metric calculations. The modulation values of the symbols associated with the feed-forward taps are hypothesized using a state trellis with A^{l_F} states. A branch metric for such a procedure may be given by:

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$$M_H(i) = \text{Re}\left\{\alpha_i^*\left[2z(i) - s_{0,i}\alpha_i - 2\sum_{l=1}^{l_F} s_{l,i}a_{i-l} - 2\sum_{l=l_F+1}^{l_{max}} s_{l,i}\hat{\alpha}_{i-l}\right]\right\},\qquad(16)$$

where $\hat{\alpha}_i$ is the tentatively demodulated symbol on the trellis path.

Similar to the MLSE embodiments described above, the number of feed-forward taps can be quantized into a finite number of values, in the extreme, to two

15   values $l_F = 0$ or L. When $l_F = 0$, the trellis reduces to one state and the receiver becomes a form of decision-feedback equalizer (DFE). In this case, the branch metric may be expressed as:

$$M_H(i) = \text{Re}\left\{\alpha_i^*\left[2z(i) - s_{0,i}\alpha_i - 2\sum_{l=l}^{l_{max}} s_{l,i}\hat{\alpha}_{i-l}\right]\right\}.\qquad(17)$$

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DFSE with an Ungerboeck metric may be improved by introducing a bias, as shown in A. Hafeez, "Trellis and Tree Search Algorithms for Equalization and Multiuser Detection," Ph.D. Thesis, University of Michigan (Ann Arbor, April 1999). Such a technique can be used with the present invention.

25   Complexity of the sequence estimator **470** may also be reduced by using a reduced-state sequence estimation (RSSE) technique along the lines proposed in M.V. Eyuboglu et al., "Reduced-State Sequence Estimation with Set Partitioning and Decision Feedback," *IEEE Trans. Commun.*, vol. COM-36, no. 1, pp. 13-20 (January

1988). According to such an approach, a set partitioning technique is used to group constellation points, which are farther apart, as a subset. An MLSE trellis is then reduced to a subset trellis in which each node represents a combination of subsets of symbols. For each transition, the symbol that has the largest branch metric is chosen

5   to represent its subset.

FIG. 7 illustrates subsets **701, 702** defined by a set partitioning scheme for a quadrature phase shift keying (QPSK) constellation **700** that can be applied in an RSSE procedure according to embodiments of the present invention. Using such a scheme, the number of trellis states can be reduced from 4^b to 2^b. FIG. 8 illustrates

10 subsets **801, 802, 803, 804** of a 16 quadrature amplitude modulation (16-QAM) constellation **800** defined under another set partitioning scheme for an RSSE procedure according to other embodiments of the invention. Using such a scheme, the number of trellis states can be reduced from 16^b to 4^b. An RSSE procedure as described above can also be combined with DFSE. According to other embodiments

15 of the invention, a state estimation procedure may be selected from a group including MLSE, DFSE, and RSSE procedures depending on l_{max}, which can be determined from the delay spread (channel estimate) and spreading factor.

According to still other embodiments of the present invention, ISI factors may be used to generate revised symbol estimates from symbol estimates generated by a

20 so-called generalized RAKE (G-RAKE) processor as described, for example, in United States Patent No. 5,572,552 to Dent et al., United States Patent Application Serial No. 09/165,647 to Bottomley, filed October 2, 1998, United States Patent Application Serial No. 09/344,898 to Bottomley et al. et. al, filed June 25, 1999 (Attorney Docket No. 8194-305), United States Patent Application Serial No.

25 09/344,899 to Wang et. al, filed June 25, 1999 (Attorney Docket No. 8194-306), and United States Patent Application Serial No. 09/420,957 to Ottosson et. al, filed October 19, 1999 (Attorney Docket No. 8194-348), each of which is incorporated herein by reference in its entirety.

For such a G-RAKE processor, the above-described initial estimate, or z-

30 parameter, may be expressed as:

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$$z(i) = \mathbf{w}^H(i)\mathbf{y}(i), \tag{18}$$

$$\mathbf{y}(i) = (y_i(iT + d_0), \ldots, y_i(iT + d_{j-1}))^T, \tag{19}$$

$$y_i(\tau) = \int_{-\infty}^{\infty} f_i^*(t)y(t + \tau)dt, \tag{20}$$

where d_j is the jth correlation time (e.g., finger delay), J is the total number of correlation times (e.g., fingers), $y_i(iT+d_j)$ is the correlator output (e.g., finger output) for correlation time d_j, and $\mathbf{w}(i)$ is the vector of combining weighting factors. It can be shown that the noise at each correlation finger output includes three components, an intersymbol interference (ISI) component, a multiuser interference (MUI) component, and a thermal noise component. It can be further shown that these noise components are statistically independent. As a result, the noise correlation between correlation fingers during the ith symbol time may be given by:

$$R(i) = R_{ISI}(i) + R_{MUI}(i) + R_N(i), \tag{21}$$

where $R_{ISI}(i)$, $R_{MUI}(i)$ and $R_N(i)$ are correlations between fingers for the ISI, MUI and thermal noise components, respectively. According to embodiments of the present invention, the weighting factors for a maximum likelihood detector, given J and $\{d_j\}_{j=0}^{J-1}$, are:

$$\mathbf{w}(i) = (R_{MUI}(i) + R_N(i))^{-1}\mathbf{h}(i), \tag{22}$$

where $\mathbf{h}(i)$ is the net channel response for symbol i. The matrix $R(i)$ accounts for noise correlation between fingers and represents knowledge of the interfering component.

In some G-RAKE receiver embodiments of the present invention, correlations to a pilot channel are performed at different lags or delays. The net channel response

h can be estimated in a number of ways. Preferably, correlations at the lags corresponding to signal rays or paths are performed. Then, using knowledge of the transmit and receive filter responses, the medium response (net response **h** minus the effects of transmit and receive filters) is determined. From the medium response, the

5   net channel response **h** may be determined by summing the contributions of the different paths using knowledge of the transmit and receive filter responses. Alternatively, the net channel response **h** can be determined by smoothing correlations at each lag. Once the net channel response **h** has been determined, the signal component on each pilot correlation may be removed, leaving instantaneous

10   noise values. These noise values may be correlated to one another and smoothed to obtain an estimate of the noise covariance **R**.

Preferably, the intersymbol interference that the equalizer will handle is not included in the noise covariance matrix **R**. To achieve this, noise values are obtained by removing all signal components handled by the equalizer from the pilot

15   correlations. The current symbol value can be removed, as normally done in a G-RAKE receiver. Intersymbol interference is removed by knowing the channel coefficient of the ISI term, as well as the cross-correlation between a current symbol spreading code and the codes used for nearby symbols that form the ISI term. The pilot symbol values are also needed if they are not the same.

20   Using a G-RAKE structure, ISI factors (*s*-parameters) analogous to the *s*-parameters described above for the conventional RAKE structure may be defined according to the relations:

$$s_{l,i} = \mathbf{w}^H(i)\mathbf{x}_{l,i} \tag{23}$$

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$$\mathbf{x}_{l,i} = (x_{l,i}(lT+d_0),\ldots,x_{l,i}(lT+d_{J-1}))^T \tag{24}$$

$$x_{l,i}(t) = f_i^*(-\tau) * h_{i-l}(\tau)$$

$$= \sum_{j=0}^{L-1} g_j \int_{-\infty}^{\infty} f_i^*(\tau) f_{i-1}(t+\tau-\tau_j)d\tau$$

$$= \sum_{j=0}^{L-1} \sum_{n=1-N}^{N-1} g_j C_{i,i-1}(n)\phi_p(t-nT_c-\tau_j). \tag{25}$$

FIG. 9 illustrates a receiver **900** according to embodiments of the present invention that uses an MLSE procedure to revise symbol estimates produced by a G-RAKE processor. An antenna **910** receives a communications signal **901**, which is

5 processed by a radio processor **920** to generate a baseband signal **925**. A G-RAKE processor **930** includes a correlator **932** that generates time-offset correlations **933** of the baseband signal **925** with a spreading sequence **945** produced by a spreading sequence generator **940**. The time-offset correlations **933** are for correlation times **937** determined by a correlation timing determiner **936** based on a channel estimate

10 **955** produced by a channel estimator **950**, for example, as described in the aforementioned United Stated Patent Application Serial No. 09/420,957 (Attorney Docket No. 8194-348).

A combiner **934** combines the time-offset correlations **933** according to weighting factors **939** generated by a weighting factor determiner **938** based on the

15 channel estimate **955**, for example, as described in the aforementioned United States Patent Application Serial No. 09/344,899 (Attorney Docket No. 8194-306). The combiner **934** produces first estimates **935** of symbols represented by the communications signal **901**. An ISI factor determiner **960** generates ISI factors **965** (*e.g.*, *s*-parameters) based on the channel estimate **955**, the spreading sequence **945**,

20 the correlation times **937** and the weighting factors **939**. A sequence estimator **970** generates second estimates **975** of the symbols from the first estimates **935** based on the ISI factors **965**. For example, as described above with reference to equation (6), the sequence estimator **970** may process the first estimates **935** according to a sequence estimation procedure that uses a branch metric that is a function of the ISI

25 factors **965**.

In a manner similar to that described above with reference to the receiver **400** of FIG. 4, the number of states used in the sequence estimator **970** may be varied responsive to the channel estimate, spreading factor, symbol modulation, and chip pulse shape function, along with the G-RAKE correlation times **937** and the

30 weighting factors **939**. For example, for some l_{max} where $s_{l_{max},i} \cong 0$, $l > l_{max}$, the

17

number of states used in the sequence estimator **970** may be $A^{l_{max}}$, where A is the number of constellation points of the symbol modulation. When the nonzero –lag s-parameters are all of small magnitudes, the sequence estimator **970** may include a symbol-by-symbol detector. In other embodiments, the value l_{max} can be quantized to

5   a finite set of values; consequently, the number of states used in the sequence estimator only take values from a finite set of integer numbers. In still other embodiments, the number of states used in the sequence estimator **970** can be either 1 or A^L, where $L > 0$ is a predetermined number. The choice of whether a one state (*i.e.* symbol-by-symbol detector) or A^L-state trellis is used in the sequence estimator may

10   be made based on the delay spread and spreading factor. For example, if the delay spread is large and the spreading factor is small, an A^L-state may be desirable. An appropriate branch metric for such a case is given by:

$$M_H(i) = \text{Re}\{\alpha_i^*[2z(i) - s_{0,i}\alpha_i]\}, \tag{26}$$

15   for the one state case, and by:

$$M_H(i) = \text{Re}\left\{\alpha_i^*\left[2z(i) - s_{0,i}\alpha_i - 2\sum_{l=1}^{L} s_{l,i}\alpha_{i-l}\right]\right\}, \tag{27}$$

20   for the A^L-state case. The aforementioned DFSE and RSSE techniques can be also applied to the G-RAKE embodiments of FIG. 9 to reduce complexity.

FIG. 10 illustrates exemplary operations **1000** for generating such s-parameters according to embodiments of the present invention. An aperiodic cross-correlation function of a spreading sequence is calculated (Block **1010**). Multiple x-

25   parameter vectors as described in equation (21) are then calculated from the aperiodic cross-correlation function of the spreading sequence, a channel estimate, and G-RAKE correlation times (Block **1020**). Inner products of the x-parameter vectors and the G-RAKE weighting factors are then determined to generate s-parameters (Block **1030**).

18

FIG. 11 illustrates an apparatus **1100,** according to still other embodiments of the present invention, for decoding a communications signal **1101** that represents a symbol sequence encoded according to a spreading sequence. A correlator **1110** generates time offset correlations **1115** of the communications signal **1101** with a

5   spreading sequence **1103.** A combiner **1120,** *e.g.,* a RAKE combiner, combines the plurality of time-offset correlations **1115** to generate first estimates **1125,** *e.g.,* decision statistics, for symbols. An estimator **1140** generates second estimates **1145** for the symbols based on ISI factors, here a plurality of weighting factors **1135** generated by weighting factor determiner circuit **1130** based on knowledge of the

10   symbol-dependence of the spreading sequence **1103,** *i.e.,* such that the weighting factors **1135** include a relationship between portions of the spreading sequence **1103.** For example, the weighting factors **1135** may be generated based on knowledge of the spreading code **1103** and a channel estimate **1102.**

The estimator **1140** may be viewed as providing a form of linear equalization.

15   The estimator **1140** includes a memory **1142,** such as a tapped delay line, that stores initial symbol estimates **1143** (*e.g.,* decision statistics) for a plurality of symbols (*e.g.,* a series of successive symbols). A combiner **1144** combines the stored initial estimates **1143** according to the weighting factors **1135** produced by the weighting factor determiner **1130** to generate revised estimates **1145** for the symbols. For

20   example, for a series of symbols S1, S2, S3, initial symbol estimates for the symbols S1, S2, S3 may be used to generate a revised estimate for symbol S2.

FIG. 12 illustrates a potential performance characteristic **1210** of a conventional receiver in comparison to a potential performance characteristic **1220** of a receiver according to embodiments of the present invention. As can be seen in FIG.

25   12, a receiver according to embodiments of the present invention may provide improved bit error rate, and more particularly, significantly improved bit error rate for higher signal to noise ratio conditions.

It will be appreciated that the present invention may be operated with multiple receive antennas, as are commonly found in cellular base stations. For such

embodiments of the present inventio, the first symbol estimates, as well as the *s*-parameters, described above may contain terms corresponding to different antennas.

In the drawings and specification, there have been disclosed typical preferred embodiments of the invention and, although specific terms are employed, they are

5   used in a generic and descriptive sense only and not for purposes of limitation, the scope of the invention being set forth in the following claims.